Transactions of the I·R·E

Professional Group on AIRBORNE ELECTRONICS

~~REP~~

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TABLE OF CONTENTS

A Report from the Chairman....K. C. Black 1

Memorandum from the Editor....John E. Wilkinson 3

Classified Symposium on Airborne Electronics.... 4

A Discussion of United Air Lines VHF Network Developments....K. J. Rhead 9

Theoretical Performance of Airborne Moving Target Indicators....Frank R. Dickey, Jr. 12

Calendar of Coming Events....Inside Back Cover

Professional Cards....Back Cover

The Institute of Radio Engineers

IRE PROFESSIONAL GROUP ON AIRBORNE ELECTRONICS

The Professional Group on Airborne Electronics is an organization, within the framework of the IRE, of members with principal professional interest in Airborne Electronics. All members of the IRE are eligible for membership in the Group and will receive all Group publications upon payment of prescribed assessments.

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Professional Group on Airborne Electronics

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A REPORT FROM THE CHAIRMAN
By K. C. Black

This article was intended originally to be a general discussion of the problems of the IRE Professional Group on Airborne Electronics. As I write this article, however, the tabulation of the voting on a change of name of the Professional Group has just been completed and I am afraid that I have to write this during a period of transition! After the proper legal machinery of the Institute is put in operation we shall have another name but, for the purposes of this article, I shall continue to refer to our organization as the Professional Group on Airborne Electronics in spite of the fact that it is not descriptive of the full scope of our activites.

Perhaps a word of explanation regarding the Professional Group structure would be worthwhile at this time. During most of our "technical" lives we have seen the Institute grow from a membership of a few thousand to over thirty thousand and I think that most of us who began our engineering work in the early thirties have felt that the Institute, although much stronger than it was in those dim days, has lost a certain amount of intimacy which used to be characteristic of the organization when one was apt to know, personally, practically every other member. The Professional Groups were created in an attempt to retain the strength of the large Institute and at the same time recapture the intimacy of a small organization. Judging from the results so far in a number of the Professional Groups it would appear that this move has been quite successful and I think that we should all give thanks to those members of the Institute who conceived the idea of the Professional Group. There is now a danger, however, that too many Professional Groups may be created and many of us are faced with the problem of belonging to so many groups that we have trouble keeping track of our correspondence. Some time ago it was agreed that, rather than create an additional Professional Group in the field of navigation, the interests of those people concerned with navigation could best be served by having them associated with the Professional Group on Airborne Electronics. The change in name which is impending is a reflection of this enlarged sphere of interest.

Although our interest is naturally great in matters pertaining to the name of our Professional Group, the health and strength of the organization depends on many things rather than the name. Our Professional Group is indeed fortunate in having a field of interest which is so important and so exciting in these days when giant strides are being made in the conquest of space. One of the most fascinating things in such a conquest is the rapidly growing ability to know where one actually is relative to other and, hopefully, fixed points on the world or in the universe. Electronics has provided a tool of unmatched usefulness in solving problems which hitherto were beyond the limits of scientific possibility. The ability to measure time and space and position has increased markedly in these last few decades.

In addition to such abilities as those mentioned in the preceding paragraph, the problem of speed is before us. Things are happening very much faster than they ever happened before. The reaction time of human beings, which formerly was able

to cope with most emergencies as well as routine operations of life, is now inadequate in many cases. Electronics has come to the assistance in this field too in that it increases by many orders of magnitude the ability of a human being to compute, to control an aircraft, or to make decisions that affect the welfare of the human being.

Looking over the entire field of interest of the Institute of Radio Engineers, I must admit that I think that we have the most exciting part assigned to us for exploration and cultivation.

The exchange of information is probably the most important single element in improving the technical ability of a group of individuals. It is of utmost importance, therefore, that we, the Professional Group, try to organize ourselves in a manner that will be most suited for such an exchange. In those parts of the country in which there is a sufficient concentration of individuals interested in the problems of aeronautical and navigational electronics, the creation of local chapters appears to be the best means of disseminating information on a week by week and month by month basis. To be sure we have our national conventions and our Professional Group is contributing more and better papers at affairs such as our March meeting in New York and August meeting in California. The Dayton Convention is now a thing of very imposing stature and is recognized throughout the country. We must not, though, overlook the fact that, in individual cities, a forum for the exchange of ideas is desirable and I believe that our principal organizing effort should now be directed toward the creation of local chapters, where the need for such a chapter is evident, and in the enthusiastic support of those chapters which have already been created. During this next season I look forward with considerable confidence to the fact that we shall have more chapters (enthusiastic ones at that!) to contribute to the basic strength of our Professional Group.

Furthermore, we must remember that the strength of the Professional Group is dependent on the energetic support of all individuals and that all of us should endeavor in whatever manner appears most practical to contribute to the welfare of the Group.

MEMORANDUM FROM THE EDITOR

John E. Wilkinson
American Phenolic Corporation

Your Editor would like to bring to your attention again the announcement of the CLASSIFIED SYMPOSIUM ON AIRBORNE ELECTRONICS, which is to be held 18 August 1953 at Univeristy of California at Los Angeles. This symposium is jointly sponsored by the IRE Professional Group on Airborne Electronics together with the Research and Development Board at Los Angeles. The following is a reprint of Headquarters letter under date of 7 July 1953, which I believe you will find self-explanatory.

We are also duplicating the program of this jointly sponsored symposium on Airborne Electronics. If you plan to attend this symposium, it would be advisable to check your clearance, for admittance will be granted only to the individuals who have been cleared.

CLASSIFIED SYMPOSIUM ON AIRBORNE ELECTRONICS

The IRE Professional Group on Airborne Electronics together with the Research and Development Board and the University of California at Los Angeles are jointly sponsoring a classified technical symposium to be held August 18, 1953 at U.C.L.A. This symposium will feature technical information normally not available because of its classified (Confidential) nature but, nevertheless, representing a major portion of the activities of the Professional Group.

The meeting is to be relatively short and informal with the presentation of papers of rather high technical content as outlined in the agenda. You are invited to attend and your participation in the discussion periods will be encouraged.

The enclosed form may be used to obtain the required clearance. Clearance takes time; time is short; no one may be admitted without clearance. Execute the clearance form at once if you wish to attend.

It has been requested that we announce that the papers to be presented do not necessarily reflect the views of the sponsoring agencies.

The attention of the Eastern engineers is called to the fact that this symposium will be held prior to the Institute of Radio Engineers West Coast Convention, which is being held in San Francisco, August 19th through August 21st. It is possible to obtain round trip, first-class tickets from the railroads or airlines which will permit you to travel by way of Los Angeles without extra cost, thereby permitting you to attend both meetings.

IRE CLASSIFIED SYMPOSIUM

AIRBORNE ELECTRONICS

Sponsored By

Los Angeles Chapter of the IRE Professional Group on Airborne Electronics, the Research and Development Board, and the University of California at Los Angeles.

MONOPULSE RADAR FOR A.I. APPLICATIONS

Dr. R. M. Ashby
Chief of the Electronics Section
Electro-Mechanical Eng. Department
North American Aviation, Inc.

A resume of the important performance characteristics of an airborne radar system will be given. In particular, typical monopulse antenna patterns, antenna impedance data, phase and gain tracking of the reference and error channels will be described.

The advantages of monopulse type tracking over the conventional conical scan tracking systems will be discussed with respect to both the tracking accuracy and relative complexity of the two systems.

A brief commentary will be made on the future possibilities of extending the performance capabilities of monopulse radar systems beyond that generally recognized at this time.

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RADAR TRACKING SYSTEMS

Dr. George F. Floyd
Member of the Technical Staff,
Radar Laboratory
Hughes Aircraft Company

After a short resume of present day radar noise theory, the effect of noise on the over-all fire control system performance will be discussed, thus allowing a comparison between the system performance to be expected with conical scan and monopulse tracking radars. In addition to the performance in the attack phase, the differences between conical scan and monopulse radars in regard to system complexity, search capabilities, anti-jamming features and other pertinent features will be mentioned. Finally, a summary of the advantages and disadvantages of various hybrid radar systems such as monocon, conopulse, etc. will be given.

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AIRBORNE RADAR SCANNERS

PART I

MECHANICAL DESIGN CONSIDERATIONS

Glenn A. Walters
Director of Research
Dalmo Victor Company

The mechanical configurations of both scanning and tracking antenna systems will be discussed and their relative merits and types of applications outlined. The characteristics of each will be analyzed on the basis of established figures of merit, i.e., volumetric efficiency, scanning efficiency, weight efficiency, etc.

PART II

SERVO DESIGN CONSIDERATIONS

Arthur P. Notthoff,
Servomechanisms Project Supervisor,
Dalmo Victor Company

Servo design problems pertaining to both stabilization and tracking systems will be discussed, outlining various methods that can be employed for application of servo power. The limitations of each design approach, their merits and over-all

problems will be compared. The following servo drives will be considered:

a. Hydraulic systems
b. Fluid magnetic clutch systems
c. Electric motor systems

The following design criteria will be discussed in connection with the above listed servo systems:

a. Maximum acceleration and velocity
b. Frequency response considerations
c. Minimum smooth tracking rate
d. Over-all system accuracy

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AN AIRBORNE RADAR SCANNER WITH HYDRAULIC DRIVES

R. B. Higley
Supervisor of Servo Unit
Radar Systems Group
North American Aviation, Inc.

The important design features for an airborne radar with hydraulic drives are discussed in detail. This includes such problems as: the gimballing of the scanner by means of a semi-circular scarf ring; cable drives, planetary gear reductions, and large angular freedom.

The pros and cons of hydraulic drives, the methods of accurate speed control with hydraulic valves, and the reasons for high-speed, high-power servos are discussed. Mention is made of the design tracking angular accuracies and servo controller components that accomplish this.

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OVER-ALL SYSTEM CONSIDERATIONS IN THE DESIGN OF A DATA LINK

Dr. Leonard C. Maier, Jr.
Electronics Laboratory
General Electric Company

The evolution of the Data Link can be used as a typical example of the requirements which must be met in designing a link which matches airborne weapon

systems to ground environments. The original change from an earlier means of modulation to that now used was dictated by estimates of the transmission accuracy, transmission rate, and number of aircraft to be controlled. From the original experimental system to the present production system, changes have been made to match many different airborne weapons, as well as many different ground environments, and further changes are possible within the present system philosophy to efficiently match the new systems which will be developed. These changes typify the problems which must be met if the Data Link is to achieve a minimum restriction of the over-all system.

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GROUND TO AIR SYNCHRONOUS BINARY DATA SYSTEM

C. A. Rypinski, Jr. and D. L. Lofquist
Collins Radio Company, Burbank, California

A communication system for transmitting binary digits from ground to aircraft has been developed and tested. The system selected operates in the military UHF communication band. Synchronous timing pulses are generated in the receiver, enabling high detection efficiency and minimum loss due to fading and other disturbances. The system has been tested with ground transmitter powers of 100 watts and 20 kilowatts. Equipment employed and test results obtained will be described.

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THE WEDDING OF THE ELECTRICAL ENGINEER

AND THE AERODYNAMICIST

Karl D. Swartzel
and
Frank Heilenday
Cornell Aeronautical Laboratories
Buffalo, New York

The science of communication has evolved several concepts which are found to be useful to the serodynamicist in the analysis of stability and control of airframes in flight. They have turned out to be of particular value in the guided missiles field where controlled guidance is accomplished without a human operator to apply ingenuity and judgement. Practical examples of the application of electrical engineering concepts are: Nyquist theorem, Bode plots, step-functional methods, super position, etc. Specific examples of the solution of the Airframe problems are demonstrated by the methods of classical aerodynamics in combination with tools borrowed from the communications field. Finally, a study is made of the impact of the concept of an

airframe as a network element in the engineering of a closed loop feedback network such as is represented by a command guidance system.

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THE CONTROLLED AIRFRAME AS AN ELEMENT OF A GUIDED MISSILE SYSTEM

R. L. Johnson
Chief, Automatic Control Systems Section
Engineering Department
Douglas Aircraft Company, Inc.

A brief description of a guided missile system is followed by the characteristics imposed upon the controlled airframe by the guidance loop of which it is a part. The effects upon airframe design of these required characteristics are covered together with the steering and roll systems envolved to control the missile. Design techniques are outlined briefly. Typical design problems encountered in the development of a satisfactory system are discussed. These include the problems of rolling moments due to steering, nonlinear pitching moments, structural vibrations coupling into the steering system, and hydraulic valve nonlinearities.

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MISSILE GUIDANCE AND CONTROL

R. W. Illman
Boeing Airplane Company

Accurate terminal guidance of an airframe requires that its control system minimize rectification of guidance noise signals. A monowing, tail-controlled airframe is inherently non-linear in response to guidance noise due to structural and performance limitations in rolling velocity and normal acceleration. Guidance and control studies of this airframe configuration have shown that performance approaching that of an ideal linear control system may be achieved by the use of a linearizing feedback and an error memory device. When guidance noise is severe, linear response is demonstrably more important than fast response in minimizing terminal error.

Guidance and control equipment must operate reliably under severe linear and angular vibrations, and certain equipments must be relatively insensitive to such vibration to minimize noise effects and to avoid parasitic oscillatory modes. Static errors (drifts) in equipment must be held small to achieve low terminal guidance error.

A DISCUSSION OF UNITED AIR LINES VHF NETWORK DEVELOPMENTS
By K. J. Rhead

The VHF network being operated by United Air Lines was designed to meet operational requirements considered desirable by the operating group. These requirements include:

A. The capability by control points located at Dispatch and C.A.A. Air Route Traffic Control Centers to talk directly with aircraft within their respective control areas as follows: any point above minimum enroute altitudes within a corridor sixty miles wide centered approximately on the flight route.

B. A "circuit busy" indication to eliminate interference resulting from attempted simultaneous use of the circuit within the interference range.

To provide coverage in the required corridor a number of radio stations over the route are interconnected for simultaneous operation so that a transmission from any control point is heard at all points. The busy signal is provided by rebroadcast of each aircraft transmission, in all areas not covered directly by the aircraft transmission.

The circuit operated between Chicago and Cleveland, employs transmitting-receiving stations at Chicago, Illinois; South Bend, Indiana; Auburn, Indiana; Toledo, Ohio; and Cleveland, Ohio. These stations are interconnected by telephone circuits to provide talking and control functions to all stations from operating points at Chicago and Cleveland.

The control circuit at Chicago or Cleveland operates and provides speech, to all transmitters approximately simultaneously. Reception of a signal at any receiver locks out the local transmitter and provides speech and carrier control to the transmitters at all other points at which a signal is not operating the associated receiver. When any transmitter is operated the associated receiver is disabled.

The line passing through the above mentioned station is approximately straight. The spacing between adjacent stations is between 75 and 90 miles. This arrangement provides coverage at all altitudes above minimum enroute in the desired corridor as well as "on the ground" coverage at Chicago, South Bend, Toledo, and Cleveland.

To avoid disturbing direct beat notes between simultaneously operated carriers on the same nominal frequency, sub-channeling about the nominal operating frequency of plus and minus ten kilocycles was tried. The stations were sub-channeled as follows: Chicago plus ten, South Bend minus ten, Auburn zero, Toledo plus ten, and Cleveland minus ten. For convenience, the above stations will be referred to by numbers 1 to 5.

With this configuration, an intermodulation problem still existed, with direct beats between stations 1 and 4 and between 2 and 5. Also, the three station

combinations 1/2/3, 2/3/4, and 3/4/5 results in the typical modulation case of $f_o = 2f_a - f_b$ which appears in the output as two ten kilocycle beats with a resultant beat between the two tens. In all of the above combinations, these beats varied about the zero point and up through the spectrum. Their magnitude was a function of the normal tolerance limits of the crystals.

It was therefore necessary to improve the stability of the transmitters before the system could be useable. An oscillator was built which limited experience indicates will hold within .00003% over reasonable periods of time. It is believed that with crystals such as are described in the IRE Proceedings (1) becoming available and techniques similar to those outlined in the Bureau of Standards Bulletin (2), that stability greatly exceeding the above can readily be obtained. The oscillator we are now using was built using only ordinary condensers and carbon resistors and it is believed that by using better quality components, it could hold frequency to .00001%.

With the tolerance of the oscillator improved, the problem of setting them to the desired frequency with the required accuracy was encountered. It was necessary to have special equipment built to accomplish this. With the oscillators set to the desired frequencies even the .00003% tolerance allowed sufficient drift to develop very low frequency beats which resulted in serious distortion of speech.

With the stability of the oscillators and with the measuring equipment available, it was then possible to sub-channel the sub-channels in various configurations in order to move the resultant beats from the very low frequency range to a band between 40 and 120 cycles. This minimized the very low frequency "chewing" of speech and kept the upper limit of beats below the pass band of the aircraft receiver. By offsetting stations 2 and 5 each plus 40 cycles (at carrier frequency of 127.5 MC) beats around the 40 and 80 cycle points were obtained but repeats of these occurred between the various combinations.

A number of various configurations have been tried with varying results, but an ideal configuration which would not repeat any combination or result in close harmonic relationship of the beats may necessitate extending the upper limit of the band used for the beat locations. This is not felt to be too serious since, as will as described later, the upper limit of the speech band has been restricted to 2500 cycles. The establishing of the lower end of the pass band to about 250/300 cycles would provide a better balance of the speech. It may also be desirable to raise the lower beat limit to 60 cycles to minimize the tolerance requirements necessary to keep the beats from approaching the undesirable range. From our experience, we feel that this undesirable range extends downward from 40 cycles.

The restriction of the speech to frequencies below 2500 cycles was made to eliminate the intermodulation effects between the upper and lower sidebands of ad-

NOTE (1) Warner, A.W., "High Frequency Crystal Units for Primary Frequency Standards", Proc IRE, Sept 52, p. 1030.

(2) "Precision Transistor Oscillator", National Bureau of Standards Tech News Bull, Feb 53, p. 17.

jacent ten kilocycle carriers which would give low frequency dissonance effects at crossover points. This cut-off point (with 2500 cycles down 20 db) was used to eliminate second harmonic components from crossing over. It may be possible to increase this upper limit when the more serious sources of intermodulation are removed and the second harmonic crossover evaluated. The upper limit of the modulation frequencies may then be raised to a point below the direct crossover of 3333 cycles.

In order to remove the direct beating of low frequency components of the speech with the low frequency beats resulting from carrier intermodulation, we have cut off the low frequency response of the transmitters at 200 cycles with the response down 30 db at 130 cycles. The characteristics of the pass band will be determined by the final intermodulation pattern.

Intermodulation effects only infrequently produce undesirable speech quality. This effect occurs only to points where a high percentage modulation condition exists between the ten kilocycle beats. This condition of course, occurs where two ten kilocycle resultants are approximately of the same level. In most cases, the impairment of speech quality due to intermodulation effects does not make it unintelligible, but rather gives it a low frequency "bubbling" sound.

We have occasionally encountered very limited areas in which the combination of all factors, including those to be discussed below, rendered the speech practically unintelligible. These areas have been so localized that normal flight time through them has been short (from a few seconds to as much as a minute) so as to make a thorough investigation. It is felt that with optimum configurations and reasonable attained frequency stability, these areas can be practically eliminated and/or the time intervals made so short as to present no serious degradation of the service.

An additional complication encountered in the network is the echo and cancellation effects. These effects occur through the speech band as a result of the delay time over the wire line between stations. It has not yet been possible to evaluate the seriousness of this factor relative to the other effects, but its existence and correction were demonstrated during a flight test made in February 1953 in cooperation with AT&T engineers.

During this test, audio checks were made and recorded using only the Chicago and Cleveland stations. The checks were made with the circuit as provided for our regular operation of the network. Tests were then duplicated with a delay loop inserted in the leg to the Cleveland transmitter. The delivery of speech to the two transmitters was brought within a time interval of a few hundred micro-seconds. The tests definitely indicated the desirability of including compensating delay networks at all transmitters.

While it is felt that the unsatisfactory speech conditions which result from carrier and sideband intermodulation effects can be brought within acceptable limits by the various means outlined, we believe that a complete investigation of receiver intermodulation characteristics for this specific case would be desirable. Any improvement which could be made in the Airborne receiver would allow a corresponding relaxation of the requirements for the ground equipment.

THEORETICAL PERFORMANCE OF AIRBORNE MOVING TARGET INDICATORS

By Frank R. Dickey, Jr.
General Electric Company

In the case of a radar moving target indicating system (MTI) in which video signals received on two successive cycles are subtracted, it is desired that signals reflected from fixed objects be completely cancelled. However, because of fluctuations in the signal strength, there is in general a residual ground clutter signal remaining after cancellation. The fluctuations may be due to slight motions of the supposedly stationary objects, to instabilities in the radar system, or to the motion of the antenna. In this paper only the effects of antenna motion are considered. That is, it is assumed that the ground is truly stationary, and that the radar system is perfectly stable. Under these conditions there is an effect due to rotation of the antenna, and, in the case of an airborne system, there will be effects due to the forward motion of the aircraft which may severely limit the performance.

The order of magnitude of most of the effects encountered has been given by Ridenour[1]. More recently, a rigorous method of calculation has been given by T. S. George[2], but his results are not in a form which is convenient for numerical calculations. The present paper describes a different method of setting up the problem which leads to simple formulas for ground clutter attenuation.

Figure 1 is an example of the type of results which can be obtained. The contours show, for a particular situation, the mean square value of residual ground clutter in decibels below the uncancelled clutter as a function of ground range and azimuth. The clutter is least, and hence the MTI performance is best along the ground track and deteriorates severely on either side.

The mean square value of the residual clutter is found to be the sum of four components, one due to rotation of the antenna, and three others due to aircraft motion. The equations for these components involve the parameters listed in Figure 1 plus the azimuth and elevation angles corresponding to positions on the radar scope. The equations make evident which ones of the system parameters are important when one wishes to design for a given performance in any particular region of the scope face.

GENERAL METHOD OF SOLUTION

In order to make calculations, a particular point on the scope, i.e., a particular range and azimuth, is selected. It is assumed that each time the antenna scans through this position an observation is made of (1) the signal voltage from the ground, and (2) the voltage remaining after subtraction of two successive signals.

[1]Ridenour, "Radar System Engineering" Rad. Lab. Series, Vol. I, McGraw Hill, 1947, pp. 657-658

[2]T. S. George, "Fluctuations of Ground Return in Airborne Radar Equipment" I.E.E. Proc., Pt. IV, Apr., 1942, v. 99, pp. 92-99

A quantity, ϵ , is then defined after a large number of such observations as follows:

$$\epsilon = \frac{\text{Mean Square Pulse-to-Pulse Voltage Change}}{\text{Mean Square Voltage}} \qquad (1)$$

The received voltage is assumed to result from a large number of small component signals arriving from various angles and randomly phased with respect to each other. It is also assumed that, on a time average, the angular distribution of energy is uniform except as influenced by certain weighting functions.

The amplitudes of the individual components are weighted, as a function of azimuth angle of arrival, according to the azimuth antenna pattern. They also are weighted as a function of elevation angle, and in the usual case where the pulse duration is short, the elevation function depends on pulse width and shape. The changes in signal strength which prevent complete cancellation are caused by (1) changes in the relative phase of various components due to displacement of the antenna platform during the interpulse period, and (2) changes in the amplitude of various components due to rotation of the antenna during the interpulse period. Although the relative phases of the various components are random, the changes in relative phase during the interpulse period can be calculated.

The differences which are observed are the differences between successive voltage envelopes since the signals are rectified by an approximately linear detector, but the problem may be simplified by considering vector differences instead of envelope differences. The vector differences are those which would result from subtraction of successive signals before rectification (and with a first order compensation for Doppler frequency shift). This mathematical artifice may be employed with little error because, if the differences are small, the mean square values of the vector differences and the envelope differences are nearly the same except for a constant factor of 1/2. Figure 2 shows the vector difference, $\mathcal{V}$, and the envelope difference, $\Delta\mathcal{V}$, at a given time. The vector difference may be divided into a component in phase and in quadrature with the sum of the two signals and, upon averaging, these components each contain half of the mean square voltage. Referring to Figure 2, the in phase component, $\mathcal{V}'$, is very nearly equal to $\Delta\mathcal{V}$ and, since $\mathcal{V}'$ has a mean square value exactly half that of $\mathcal{V}$, one can say that $\Delta\mathcal{V}$ has a mean square value approximately half that of $\mathcal{V}$.

Assume that one has available the following functions:

$A(\theta)$ = round-trip voltage antenna pattern in azimuth,

$B(\phi)$ = elevation weighting function usually determined by pulse shape,

$F(\theta,\phi)$ = change in relative phase produced by a given displacement of the antenna,

θ, ϕ = azimuth and elevation angles, respectively, of individual signal components, measured with respect to the center of the effective beam.

The individual voltage components comprising a signal are weighted according to $A(\theta)\,B(\phi)$. The vector difference between two successive signals is composed of the same voltage components weighted according to the following function:

$$A(\theta + \tfrac{1}{2}\omega_a T)\,B(\phi)\,e^{+\frac{1}{2}jF(\theta,\phi)} - A(\theta - \tfrac{1}{2}\omega_a T)\,B(\phi)\,e^{-\frac{1}{2}jF(\theta,\phi)} \qquad (2)$$

where

ω_a = angular rotation rate of antenna (about a vertical axis),

T = time interval between pulses.

With the assumptions regarding the statistical nature of the signal components, the ratio of mean square voltages is:

$$\epsilon = \frac{\frac{1}{2}\int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty}\left| A(\theta + \tfrac{1}{2}\omega T)\,B(\phi)\,e^{+\frac{1}{2}jF(\theta,\phi)} - A(\theta - \tfrac{1}{2}\omega_a T)\,B(\phi)\,e^{-\frac{1}{2}jF(\theta,\phi)}\right|^2 d\theta\, d\phi}{\int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty}\left| A(\theta)\,B(\phi)\right|^2 d\theta\, d\phi} \qquad (3)$$

In order to put this expression into a more useful form, a rectangular coordinate system is oriented in such a way that the z axis is along the center of the effective beam, and the x axis is horizontal. The y axis in general is not vertical, but, as shown in Figure 3, it falls in the same vertical plane as the z axis. The displacement of the antenna during the interpulse period is expressed as a component along each axis. These components are:

$$X = vT \sin(\Theta - \Theta_0) \qquad (4)$$

$$Y = vT \cos(\Theta - \Theta_0)\sin(\Phi - \Phi_0) \qquad (5)$$

$$Z = vT \cos(\Theta - \Theta_0)\cos(\Phi - \Phi_0) \qquad (6)$$

where v = velocity of the aircraft,

T = time interval between pulses

Θ, Φ = azimuth and elevation angles respectively of the z axis or center of the effective beam,

Θ_o, Φ_o = azimuth and elevation angles respectively of the direction of motion.

The angles Θ and Φ_o are measured from some arbitrary azimuth angle, for example, the heading of the aircraft. The angles Φ and Φ_o are measured downward from the horizon. For horizontal motion Φ_o is zero. The value of Φ is:

$$\Phi = \tan^{-1} \frac{h}{G} \tag{7}$$

where h = altitude

G = ground range corresponding to the chosen point on the radar scope.

The function $F(\theta, \phi)$ which expresses the phase shift produced by a given antenna displacement is:

$$F(\theta,\phi) = \frac{4\pi}{\lambda}\left[X \sin\theta + Y \cos\theta \sin\phi + Z(\cos\theta \cos\phi - 1)\right] \tag{8}$$

where λ is the wavelength, θ and ϕ are the angles of arrival with respect to the Z axis as previously defined and X, Y, and Z are the displacements as described above. It is assumed that all reflecting objects are at a great distance from the antenna.

If it is assumed that the beam widths are small, Equation 8 may be replaced by the following approximation:

$$F(\theta,\phi) = \frac{4\pi}{\lambda}\left[X\theta + Y\phi - Z(\theta^2 + \phi^2)\right] \tag{9}$$

The Z term is usually negligibly small, but is retained in order to take care of cases in which the coordinate system is oriented in such a way that X and Y are zero. This occurs at long range along the ground track.

When Equation 9 is substituted into Equation 3, with the provision that the beam widths are narrow, Equation 3 may be separated approximately into four terms as follows:

$$\epsilon = \epsilon_R + \epsilon_X + \epsilon_Y + \epsilon_Z \tag{10}$$

where

$$\epsilon_R = \frac{\frac{1}{2}\int_{-\infty}^{+\infty} \left|A(\theta + \frac{1}{2}\omega_a T) - A(\theta - \frac{1}{2}\omega_a T)\right|^2 d\theta}{\int_{-\infty}^{+\infty} |A(\theta)|^2 d\theta} \tag{11}$$

$$\epsilon_X = \frac{2\int_{-\infty}^{+\infty} |A(\theta)|^2 \sin^2\left(\frac{2\pi X \theta}{\lambda}\right) d\theta}{\int_{-\infty}^{+\infty} |A(\theta)|^2 d\theta} \tag{12}$$

$$\epsilon_Y = \frac{2\int_{-\infty}^{+\infty} |B(\phi)|^2 \sin^2\left(\frac{2\pi Y \phi}{\lambda}\right) d\phi}{\int_{-\infty}^{+\infty} |B(\phi)|^2 d\phi} \tag{13}$$

$$\epsilon_Z = \frac{2\int_{-\infty}^{-\infty}\int_{-\infty}^{+\infty} |A(\theta)\, B(\phi)|^2 \sin^2\left(\frac{2\pi Z}{\lambda}(\theta^2 + \phi^2)\right) d\theta\, d\phi}{\int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} |A(\theta)\, B(\phi)|^2 d\theta\, d\phi} \tag{14}$$

These are the same expressions which one obtains by specializing Equation 3, first to rotation only, then to X displacement only, etc. Thus, to a first approximation, the various components of motion contribute independently to the mean square value of uncancelled ground clutter.

By making use of Parseval's theorem[3], the expressions for ϵ_R, ϵ_X, and ϵ_Y, may be written in alternative forms. Alternative expressions for ϵ_R and ϵ_X, are:

$$\epsilon_R = \frac{2\int_{-\infty}^{+\infty} |F(\beta)|^2 \sin^2(\pi \omega_a T \beta)\, d\beta}{\int_{-\infty}^{+\infty} |F(\beta)|^2 d\beta} \tag{15}$$

$$\epsilon_X = \frac{\frac{1}{2}\int_{-\infty}^{+\infty} \left|F(\beta - \frac{X}{\lambda}) - F(\beta + \frac{X}{\lambda})\right|^2 d\beta}{\int_{-\infty}^{+\infty} |F(\beta)|^2 d\beta} \tag{16}$$

[3] The required identity may be found in Campbell and Foster, "Fourier Integrals for Practical Applications", D. Van Nostrand Company, 1948, Footnote page 39.

where $F(\beta)$ is the Fourier transform of $A(\theta)$.

Instead of starting with the antenna patterns, one can equally well start with an assumed aperture illumination function. $F(\beta)$ may be obtained by convolution of the aperture function as follows:

$$F(\beta) = \int_{-\infty}^{+\infty} G(\gamma)\, G(\beta-\gamma)\, d\gamma \tag{17}$$

where G = function describing aperture field strength,

γ = horizontal distance from center of the aperture in terms of wavelengths.

The ϵ_R term describes the effect of scanning while the ϵ_x term describes the effect of motion perpendicular to the center of the antenna beam. It may be noted that there is a symmetrical relationship between the scanning and the motion effects. This symmetry may be made more evident by the following change of variables. In the expressions for ϵ_R let:

$$\theta = \omega_a t \tag{18}$$

$$\beta = \frac{f}{\omega_a} \tag{19}$$

In the expressions for ϵ_x, let:

$$\beta = \frac{2 v_x t}{\lambda} = \frac{2 v t \sin(\theta_1 - \theta_0)}{\lambda} \tag{20}$$

$$\theta = \frac{\lambda f}{2 v_x} = \frac{\lambda f}{2 \cdot v \sin(\theta_1 - \theta_0)} \tag{21}$$

The variable t may be interpreted physically as time, and the variable f may be interpreted physically as frequency. Then as functions of time, the expressions are:

$$\epsilon_R = \frac{\frac{1}{2}\int_{-\infty}^{+\infty} \left| A\left(\omega_a\left(t+\frac{T}{2}\right)\right) - A\left(\omega_a\left(t-\frac{T}{2}\right)\right)\right|^2 dt}{\int_{-\infty}^{+\infty} \left| A(\omega_a t)\right|^2 dt} \tag{22}$$

$$\epsilon_x = \frac{\frac{1}{2}\int_{-\infty}^{+\infty}\left|F\left(\frac{2v_x}{\lambda}\left(t-\frac{T}{2}\right)\right)-F\left(\frac{2v_x}{\lambda}\left(t+\frac{T}{2}\right)\right)\right|^2 dt}{\int_{-\infty}^{+\infty}\left|F\left(\frac{2v_x t}{\lambda}\right)\right|^2 dt} \quad (23)$$

As functions of frequency, the expressions are:

$$\epsilon_R = \frac{2\int_{-\infty}^{+\infty}\left|F\left(\frac{f}{\omega_a}\right)\right|^2 \sin^2(\pi T f)\, df}{\int_{-\infty}^{+\infty}\left|F\left(\frac{f}{\omega_a}\right)\right|^2 df} \quad (24)$$

$$\epsilon_x = \frac{2\int_{-\infty}^{+\infty}\left|A\left(\frac{\lambda f}{2v_x}\right)\right|^2 \sin^2(\pi T f)\, df}{\int_{-\infty}^{+\infty}\left|A\left(\frac{\lambda f}{2v}\right)\right|^2 df} \quad (25)$$

The functions which are integrated in a typical case are illustrated in Figures 4 and 5. Using time as a variable, one has in the scanning case the difference between two displaced antenna patterns, while in the motion case one has the difference between two displaced aperture functions. Using frequency as the variable, one has in each case a sine function which represents the effect of delaying one signal and subtracting it from another, and another function which represents the spectrum of the signal fluctuations. In the motion case the spectrum has the shape of the antenna pattern, while in the scanning case the spectrum has the shape of the transform of the antenna pattern. It is of interest to note that for any aperture of finite width, a, the frequency spectrum for scanning fluctuations lies entirely below a given frequency,

$$f_{max} = \frac{a\,\omega_a}{\lambda} \quad (26)$$

RESULTS

Expressions for ϵ_R, ϵ_X, and ϵ_Y have been worked out assuming the following azimuth antenna patterns (round trip):

Rectangular $$F(\theta) = \begin{cases} 0 & \theta < -\frac{1}{2}\Delta\theta \\ 1 & -\frac{1}{2}\Delta\theta < \theta < +\frac{1}{2}\Delta\theta \\ 0 & +\frac{1}{2}\Delta\theta < \theta \end{cases} \tag{27}$$

Gaussian $$F(\theta) = e^{-2\left(1.777\frac{\theta}{\Delta\theta}\right)^2} \tag{28}$$

Sin x/x $$F(\theta) = \left[\frac{\sin\left(2.78\frac{\theta}{\Delta\theta}\right)}{\left(2.78\frac{\theta}{\Delta\theta}\right)}\right]^2 \tag{29}$$

where $\Delta\theta$ = azimuth beam width measured between half power points on the one way pattern.

These functions may also be used in the elevation direction. However, since the elevation weighting function depends on pulse shape, the rectangular and the Gaussian functions may be suitable, but the sin x/x function is not appropriate. The effective elevation beamwidth is given approximately by

$$\Delta\phi = \frac{\frac{1}{2}C\tau \tan\Phi \sin\Phi}{h} \tag{30}$$

where $\Delta\phi$ = effective elevation beamwidth

C = velocity of light

τ = pulse duration

h = altitude

Φ = depression angle as previously defined.

The following symbols are used:

$$N = \frac{\Delta\theta}{\omega_a T} \tag{31}$$

$$M_X = \frac{\pi X \Delta\theta}{\lambda} = \frac{\pi v T \Delta\theta \sin\theta}{\lambda} \tag{32}$$

$$M_Y = \frac{\pi Y \Delta\phi}{\lambda} = \frac{\pi v c T \tau \cos\theta \sin^2\phi \tan\phi}{\lambda h} \tag{33}$$

Equations for ϵ_R

Rectangular Pattern $\epsilon_R = \frac{1}{N}$ (34)

Gaussian Pattern (35)

$$\epsilon_R = 1 - e^{-\left(\frac{1.177}{N}\right)^2} = \left(\frac{1.177}{N}\right)^2 - \frac{1}{2!}\left(\frac{1.177}{N}\right)^4 + \frac{1}{3!}\left(\frac{1.177}{N}\right)^6 - \cdots$$

Sin x/x Pattern

$$\epsilon_R = 1 - 6\left(\frac{N}{2\cdot 2.78}\right)^2 \left(1 - \frac{\sin\left(\frac{2\cdot 2.78}{N}\right)}{\left(\frac{2\cdot 2.78}{N}\right)}\right) \tag{36}$$

$$= 6\left[\frac{1}{5!}\left(\frac{2\cdot 2.78}{N}\right)^2 - \frac{1}{7!}\left(\frac{2\cdot 2.78}{N}\right)^4 + \frac{1}{9!}\left(\frac{2\cdot 2.78}{N}\right)^6 - \cdots\right] \tag{37}$$

Equations for ϵ_x

Rectangular Pattern (38)

$$\epsilon_x = 1 - \frac{\sin 2M_x}{2M_x} = \frac{1}{3!}(2M_x)^2 - \frac{1}{5!}(2M_x)^4 + \frac{1}{7!}(2M_x)^6 - \cdots$$

Gaussian Pattern (39)

$$\epsilon_x = 1 - e^{-\left(\frac{M_x}{1.177}\right)^2} = \left(\frac{M_x}{1.177}\right)^2 - \frac{1}{2!}\left(\frac{M_x}{1.177}\right)^4 + \frac{1}{3!}\left(\frac{M_x}{1.177}\right)^6 - \cdots$$

Sin x/x Pattern

$$\epsilon_x = \begin{cases} 6\left(\frac{M_x}{2.78}\right)^2\left(1 - \frac{|M_x|}{2.78}\right) & 0 < \frac{|M_x|}{2.78} < \frac{1}{2} \quad (40) \\ 1 - 2\left(1 - \frac{|M_x|}{2.78}\right)^3 & \frac{1}{2} < \frac{|M_x|}{2.78} < 1 \quad (41) \\ 1 & 1 < \frac{|M_x|}{2.78} \quad (42) \end{cases}$$

$$\epsilon_x = \begin{cases} 6\left(\frac{X}{a}\right)^2\left(1-\frac{|X|}{a}\right) & 0 < \frac{|X|}{a} < \frac{1}{2} \quad (43) \\ 1-2\left(1-\frac{|X|}{a}\right)^3 & \frac{1}{2} < \frac{|X|}{a} < 1 \quad (44) \\ 1 & 1 < \frac{|X|}{a} \quad (45) \end{cases}$$

where a = width of uniformly illuminated aperture.

Equations for ϵ_Y are the same as those for ϵ_X except that M_x is replaced by M_y. Also the sin x/x case is not appropriate for ϵ_Y

The last component, ϵ_Z, is usually negligible. Its value assuming a rectangular pattern is:

$$\epsilon_z = \frac{1}{10}\left(\frac{\pi Z (\Delta\theta)^2}{\lambda}\right)^2 - \frac{1}{64 \cdot 54}\left(\frac{\pi Z (\Delta\theta)^2}{\lambda}\right)^4 + \cdots \quad (46)$$

The power series expressions are more convenient for numerical calculations than the closed form expressions. Usually the first term of the series will give the result with sufficient accuracy. It may be noted that, for the most part, the coefficients are roughly the same regardless of the assumed shape of the beam.

At any point on the scope, some one of the components is likely to be larger than the others. Figure 6 indicates the regions where each component is likely to predominate and also lists typical first terms in a form which shows the angular dependence. The X term is important in the regions to the side. The Y component becomes of importance along the ground track where the X component goes to zero. It is large however only where the depression angle is large and is therefore important at high altitudes. The scanning component is dependent on the rate of rotation of the antenna and is independent of the scope position. This component may limit the performance at long range and along the ground track where the X and Y components both become small. In this region also, the Z component may be appreciable.

It is of interest that the X component, which generally represents the largest fluctuations in the case of airborne radar, can be expressed as a function of x displacement divided by aperture width. For example, using Equation 43, one finds that to attenuate ground clutter at 90 degrees from the ground track by 20 decibels, the distance travelled by the aircraft between pulses must not be greater than 4% of the antenna width. This is true regardless of the wavelength or beamwidth.

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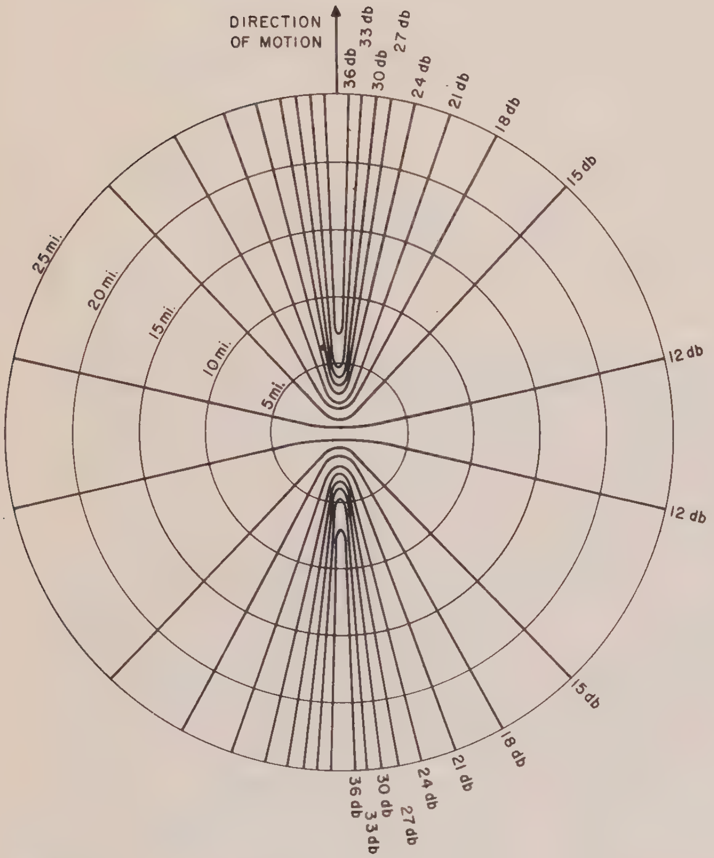


BEAM WIDTH	3.0 DEGREES	PULSE LENGTH	3/4 μSEC.
GROUND SPEED	250 KNOTS	PRF	2000
ALTITUDE	20,000 FEET	WAVELENGTH	3.2 CM.
ANTENNA ROTATION 12 RPM			

FIG. 1 TYPICAL GROUND CLUTTER ATTENUATION FOR AIRBORNE MTI

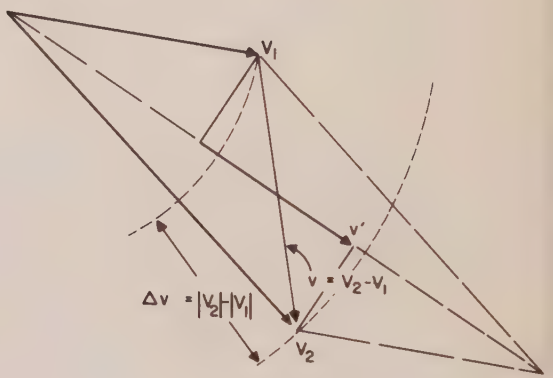


FIG. 2

VECTOR DIAGRAM SHOWING NEAR EQUALITY BETWEEN LENGTHS OF $|V_2|-|V_1|$ AND THE COMPONENT OF $V_1 - V_2$ IN PHASE WITH $V_1 + V_2$

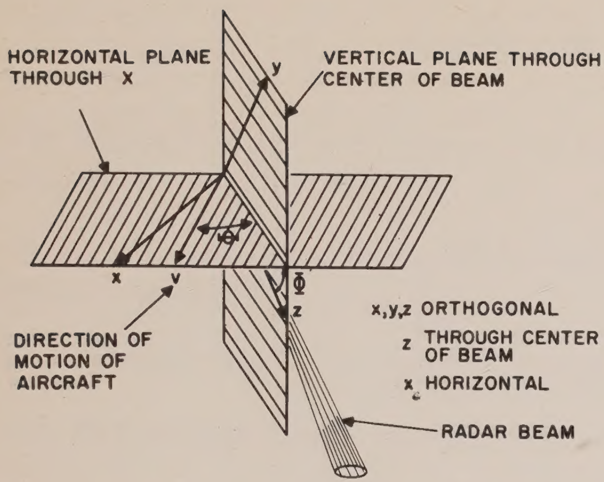


FIG. 3 – LOCATION OF COORDINATE AXES

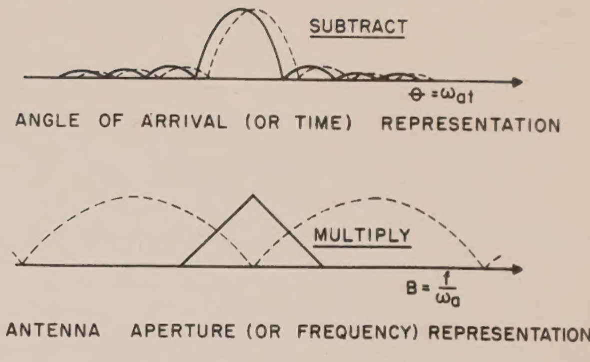


FIG. 4 SCANNING EFFECTS

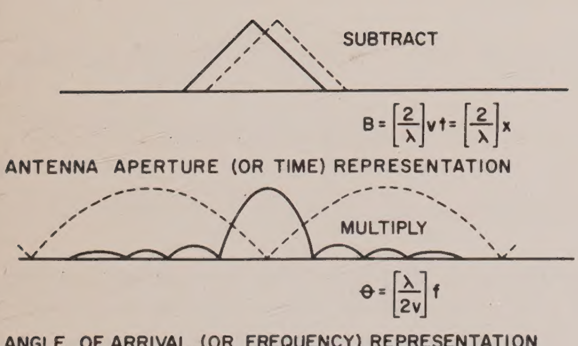
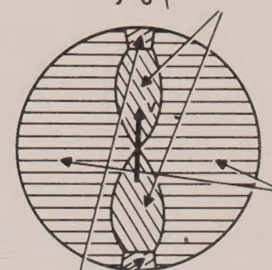


FIG. 5 MOTION EFFECTS

$$\epsilon = \epsilon_x + \epsilon_y + \epsilon_z + \epsilon_R$$

$$\epsilon_y = \frac{1}{6}\left(\frac{\pi v c T \tau \cos\Theta \sin^2\Phi \tan\Phi}{h\lambda}\right)^2$$



$$\epsilon_x = 0.78\left(\frac{\pi v T(\Delta\Theta)\sin\Theta}{\lambda}\right)^2 = 6\left(\frac{vT \sin\Theta}{a}\right)^2$$

$$\epsilon_R = 1.54\left(\frac{\omega_a T}{\Delta\Theta}\right)^2$$

OR $$\epsilon_z = \frac{1}{10}\left(\frac{\pi v T(\Delta\Theta)^2 \cos\Theta \cos\Phi}{\lambda}\right)^2$$

FIG. 6 - COMPONENTS OF MEAN SQUARE RESIDUAL GROUND CLUTTER AND REGIONS WHERE EACH COMPONENT IS LIKELY TO BE LARGEST

CALENDAR OF COMING EVENTS

1953 Western Electronics Show and Convention, Civic Auditorium, San Francisco, California, August 19-21

Eighth National Instrument Conference and Exhibit, Hotel Sherman, Chicago, Illinois, September 21-25

National Electronic Conference, Hotel Sherman, Chicago, Illinois, September 28-30

1953 IRE-RTMA Radio Fall Meeting, Toronto, Ontario, October 26-28

Sixth Southwestern IRE Conference and Electronics Show, Tulsa, Oklahoma, February 4-6

Radio Engineering Show and IRE National Convention, Hotel Waldorf-Astoria, New York City, March 22-25

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